DESIGN OF A CARRIER-TELEPHONE TRANSLATOR

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THESIS

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ABSTRACT

A design proposal for a ten-channel carrier-telephone translator is presented. The lower-sideband portion was built and tested in the laboratory and the results are reported. The theory of operation and tuning procedure are discussed. A brief history of multiplex telephony is also presented. The translator is designed to interface with a microwave radio link.

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I. INTRODUCTION

On 8 August 1969 the Federal Communications Commission authorized Microwave Communications Inc. to offer microwave service between St. Louis and Chicago in competition with regulated communications carriers such as American Telephone & Telegraph and General Telephone & Electronics.

The FCC decision marked the first time the agency has authorized anyone other than regulated carriers to offer private-line microwave service.

As a result of the FCC decision, it is anticipated that there will be a growing demand for low-capacity microwave service by small businesses that want a private communications link between their various plants.

This thesis is a design proposal for a translator for a ten-channel telephone system which would interface with an existing microwave link. It is hoped that the translator would be competitive with existing equipment in cost as well as quality.

II. BRIEF HISTORY OF MULTIPLEX TELEPHONY

The first telephone capable of practical use was invented by Alexander Graham Bell in 1876. The first telephone circuits were extremely crude, consisting of a single grounded wire with a telephone connected at each end. With this arrangement, each telephone could be connected only with the telephone at the opposite end of the circuit - and not to any others. Such a simple arrangement was very limited.

By 1878 switching mechanisms had been developed whereby any two telephones in a large system could be connected together. The first commercial telephone office opened on January 28, 1878 in New Haven, Connecticut serving twenty-one telephones over eight open-wire lines called subscriber loops /1/. Service was soon extended by interconnecting the switchboards in the telephone offices with additional wire lines which became known as trunks. Trunks interconnecting local offices were designated exchange trunks, while those inter connecting long-distance offices were designated toll trunks.

The telephone industry rapidly expanded until soon there were hundreds of wire lines, carried on crossarms mounted on wooden poles, appearing along all major streets and roads throughout the country.

It was soon learned that the single-wire line with ground return, borrowed from the telegraph industry, was not completely suitable for telephone communications because of excessive noise and electrical disturbances that were annoying to the user. This problem was solved with the development of the two-wire, or metallic circuit. This type of circuit consisted of two closely paralleled wires, with one of the wires providing the current return path instead of returning the current through the earth.

Although the metallic circuit solved an interference problem, it created the problems of reconstructing practically the entire telephone plant and doubling the already burdensome, and oftentimes unsightly, mass of wire lines. This enormous task was performed by the telephone industry during the period 1890 - 1900.

Putting the wire pairs into cables served to remove some of the wire lines from view, but the problem of continually having to enlarge the outside wire plant to satisfy the increasing demand for more circuits still remained. A method of increasing the number of telephone circuits without having to add thousands of miles of more wire was sorely needed.

Early in the development of telephone communications, it was found that a frequency range from about 300 to 2800 Hz would convey speech with sufficient fidelity and clarity for commercial telephone service. (Modern telephone systems transmit speech signals ranging from about 300 to 3400 Hz) Since it was possible to transmit frequencies up to hundreds of kHz over wire lines, not all of the capabilities of the existing telephone circuits were being used. This fact resulted in a search for a means of transmitting more than one telephone conversation simultaneously over a single pair of wires, a process known as multiplexing.

The principles of multiplexing had already been developed by the telegraph industry, but a practical multiplexer for voice telephony had to await the invention of the vacuum-tube triode and the electrical filter.

In order to transmit two or more telephone signals simultaneously over the same circuit, the signals must be separated so that they do not interfere with each other. This can be done by separating them either in frequency, known as frequency-division multiplexing, or in time, known as time-division multiplexing. Both methods were tried experimentally in the early stages of multiplex telephone; however, frequency-division multiplexing became the more prevalent.

Either frequency modulation or amplitude modulation may be used to transform speech signals to separate frequency bands, as required for frequency-division multiplexing. Amplitude modulation is most commonly used. In this type of modulation, the resulting modulated wave consists of a carrier wave, an upper sideband wave, and a lower sideband wave. The two sideband waves are separated from the carrier wave by a frequency equal to the modulating speech signal. Each sideband wave contains all of the frequency components of the modulating speech signal. It soon became evident that only one sideband wave had to be transmitted - provided an equivalent carrier was available at the receiving end to demodulate the signal. By using only one sideband, the energy required to transmit the signal is reduced considerably and the frequency band used is essentially half of that required if both sidebands and the carrier are transmitted. Thus, twice as many telephone channels can be transmitted in the same multiplex frequency band. The technique of transmitting only one sideband, known as sigle-sideband suppressed-carrier, is used in most of the multiplex systems that have been developed for toll circuit use. The underlying principle whereby the voice frequencies modulate a higher frequency current which "carries" the voice currents led to the name "carrier telephone system".

Following experiments between Toledo, Ohio and South Bend, Indiana, in 1917, the first commercial application of the carrier principle was made in 1918 on an open-wire line between Baltimore, Maryland and Pittsburgh, Pennsylvania, giving four additional telephone circuits on a pair of wires. [2]

The Bell System designated their original multiplex system as type A, thus beginning a succession of different systems with alphabet designations.

The development of extremely linear electronic amplifiers using negative feedback techniques, along with other technical advances, led to the development of a twelve-channel system designated type J. The first type J system was installed between Toledo and South Bend in 1937. The type J system was designed for open-wire lines and had a line frequency range of 36 to 140 kHz. The frequency band from 36 to 84 kHz was used for transmission in one direction and the frequency band from 92 to 140 kHz was used for transmission in the opposite direction.

The next twelve-channel system, designated type K, was used on a transcontinental cable system which was put into service in 1938. This system used the frequency band from 12 to 60 kHz for transmission in both directions. This was done by using a different wire pair for each direction, thus establishing a physical four-wire system.

The lower line frequency of the type K system was achieved using a new technique called group modulation. In the earlier systems, the multiplex line frequencies were accomplished by a single direct-modulation step. Group modulation, however, consists of using two or more steps of modulation to establish the line frequencies. One of the most significant advantages of group modulation was that it provided a simplified means of interconnecting standard subgroups of channels at line frequencies, a technique employed extensively in later multiplex systems.

Once multiplexing had been firmly established as the standard method of deriving toll circuits, there was increasing pressure to reduce the physical size of the equipment. During and following World War II, a continuing effort was made to reduce the size of electronic components, and to add new devices, such as germanium and silicon diodes and transistors, which were small and required much less power than vacuum tubes. This led to a standard 24-channel miniaturized multiplex system that became economical for short-haul applications.

To compete with the cost of loaded cable pairs used for exchange trunks shorter than 10 miles, the Bell System developed a multiplex system significantly different from the conventional frequency-division systems used in the past. This system, designated type T1, was a time-division multiplex system which used pulse-code modulation to provide 24 telephone channels over an exchange trunk cable.

The 24-channel systems were near the capacity limits for the existing transmission mediums. However, multiplex systems could be made to operate with much greater bandwidths than those provided by conventional wire systems. The needed wideband transmission medium could be provided by coaxial cables or microwave radio systems, and the wide introduction of television made it economical to install these facilities. In about 1948, the Bell System completed a transcontinental coaxial cable transmission facility, designated type L1, to be used for television as well as to provide a large number of telephone channels. The L1 facility was capable of handling up to 600 single-sideband suppressed-carrier frequency-division multiplex telephone channels, or one television channel. Later, a higher capacity coaxial cable, designated type L3, was developed which was capable of handling 1,860 multiplex telephone channels, or one television channel and 600 multiplex telephone channels. The L3 coaxial cable was followed by the L4 coaxial cable which had a capacity of 32,000 multiplex telephone channels.

A microwave radio-relay system was placed in service between New York City and Boston in 1947, and one across the United States in 1951. The Bell System type TH microwave system, operating in the 6000-MHz common-carrier band provided a capacity of over 11,000 multiplex channels.

The Bell System developed a special single-sideband suppressed-carrier frequency-division multiplex system for use with the high-capacity coaxial cable and microwave links. The multiplex system was designated type L and could provide up to 600, or alternatively, up to 1,860 multiplex telephone channels.

The type L multiplex system combined the voice-frequency telephone circuits into 12-channel groups. A series of group-modulation steps were then used to form up to fifty 12-channel groups (600 channels) into a baseband with a frequency range of 60 to 2788 kHz, or up to 155 12-channel groups (1860 channels) into a baseband with a frequency range of 312 to 8284 kHz.

The telephone industry has grown significantly since 1876, when Alexander Graham Bell was able to call Mr. Watson from the next room, and July 20, 1969 when millions of people on earth looked on while President Richard M. Nixon carried on a radiotelephone conversation with astronauts Armstrong and Aldrin who were standing on the moon.

Waveguide transmission systems and laser-beam transmissions offer the possibility of wideband systems with capacities up to a million, or more, multiplexed telephone channels.

III. SPECIFICATIONS AND DESIGN CONSIDERATIONS

A. SPECIFICATIONS

The specifications listed below are required by the equipments with which the translator is designed to interface.

Method of Transmission	Frequency-division, single-sideband, suppressed-carrier
Channel Capacity	1 to 10 channels
Input/Output Impedance	75 ohms
Modulator Input Level	-45 dbm nominal per channel
Modulator Output Level	-15 dbm nominal per channel
Demodulator Input Level	-15 dbm nominal per channel
Demodulator Output Level	-15 dbm nominal per channel
Modulator Input Frequency	4-24 kHz
Modulator Output/Demodulator Input Frequency	28-48 kHz and 56-76 kHz with a -25 dbm 52-kHz carrier pilot
Demodulator Output Frequency	4-24 kHz
Carrier Frequency	52 kHz
Maximum Ripple	1 db
Power Supply Voltage Current Drain (actual)	Negative 24 volts dc 80 ma.
Synchronous System	Yes, must be able to handle digital data transmissions

B. DESIGN CONSIDERATIONS

The prime consideration in the design of the translator was cost. Most of the cost of such a system is fixed by the filters; however, the cost was kept in mind during the design. For example, the number of different sizes of resistors and capacitors was kept as small as possible in order to take advantage of the savings possible when ordering in large quantities. Integrated circuits were used to save on components and assembly time. The integrated circuits also add to the reliability and quality of the system.

A synchronous system is required in order to meet the specification that the system be able to handle digital information such as teletype and computer outputs.

Standard carrier-telegraph systems multiplex up to 20 or more telegraph circuits into one standard 4-kHz voice channel employing frequency-division techniques. The most prevalent form of modulation is frequency-shift keying (FSK); however, amplitude modulation can also be used.

The channel spacing for FSK systems is commonly 120 or 170 Hertz. The 120-Hertz spacing is used for systems operating at 60 words per minute or less, while 170-Hertz spacing is used for 100 words per minute or less. The usual frequency shift for 120-Hertz-spaced channels is $\pm$ 30 Hertz and the frequency shift for 170-Hertz-spaced channels is $\pm$ 42.5 Hertz. [3]

For voice transmission the reinserted carrier at the demodulator can be several Hertz different from the original carrier without seriously affecting the quality of the speech; however, the difference in carrier frequencies must be kept very small for digital data transmissions.

In a synchronous system, the carrier frequency at the demodulator is exactly the same as that used at the modulator; hence, the error is zero.

IV. THEORY OF OPERATION

A. GENERAL

A block diagram of the proposed system is shown in Figure 1. Each modulator takes five telephone voice channels and translates them to a position in the frequency spectrum which is suitable to modulate a microwave transmitter. First, consider channels 1 through 5. Each channel is 4-kHz wide and all have previously been frequency multiplexed such that each channel fills a 4-kHz slot in the frequency spectrum 4-24 kHz, starting with channel 1 in the 4-8 kHz slot and ending with channel 5 in the 20-24 kHz slot. The output of the balanced modulator contains the upper and lower sidebands with the carrier being suppressed. The upper sideband is selected by the high-pass filter.

Channels 6-10 are identical to channels 1-5, but in this case the lower sideband is selected by the low-pass filter.

The two sidebands are combined to form a baseband suitable for modulating a transmitter. A controlled amount of carrier signal is inserted to synchronize the demodulator.

At the receiving end , the two sidebands are separated by filters and applied to a balanced demodulator. The carrier is also picked off through a narrow-bandpass crystal filter and amplified to provide the oscillator signal to drive the demodulators. The output of the demodulator contains the desired frequencies plus an unwanted upper sideband which is filtered out by the low-pass filters.

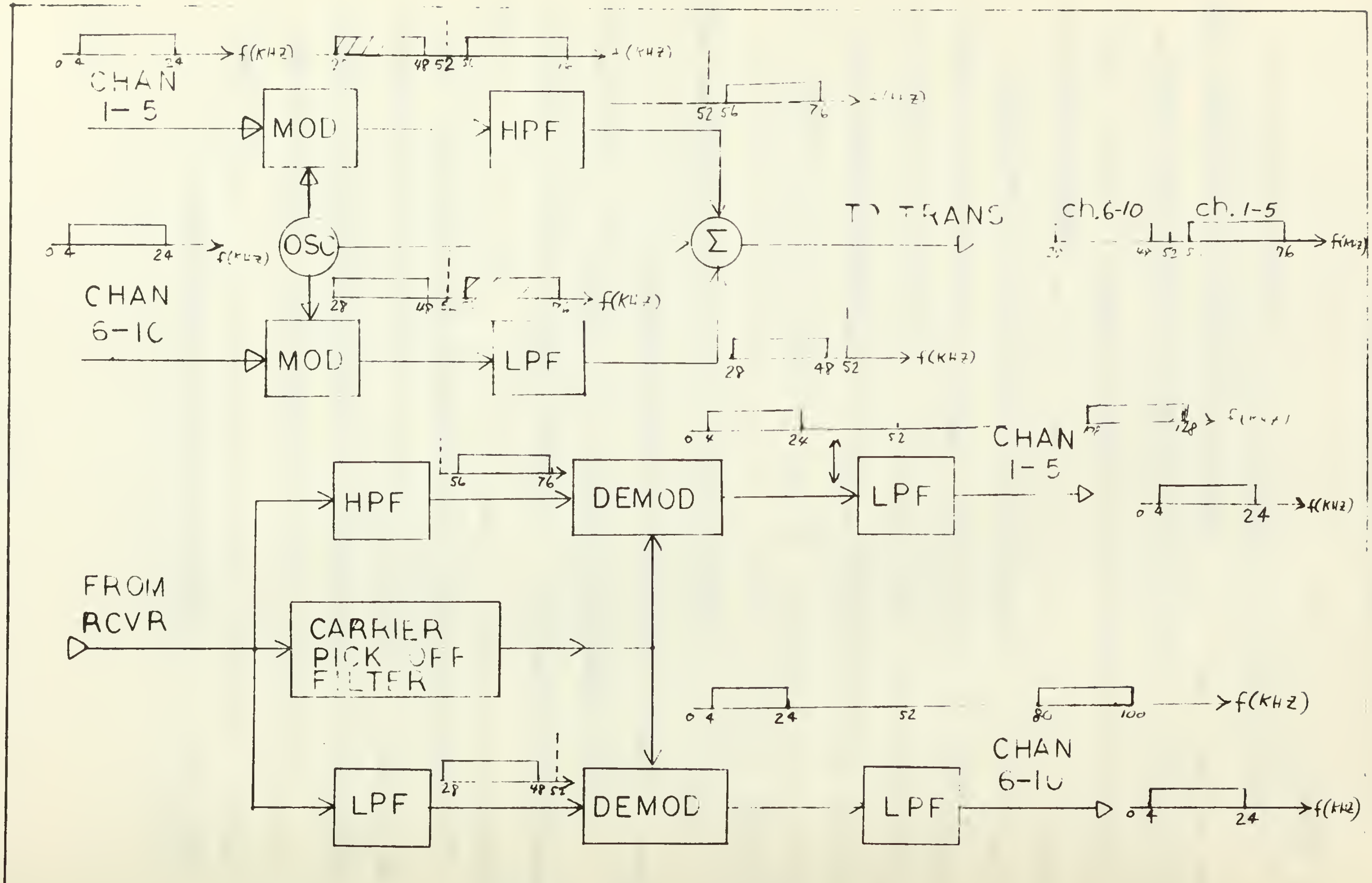


Figure 1. System Block Diagram

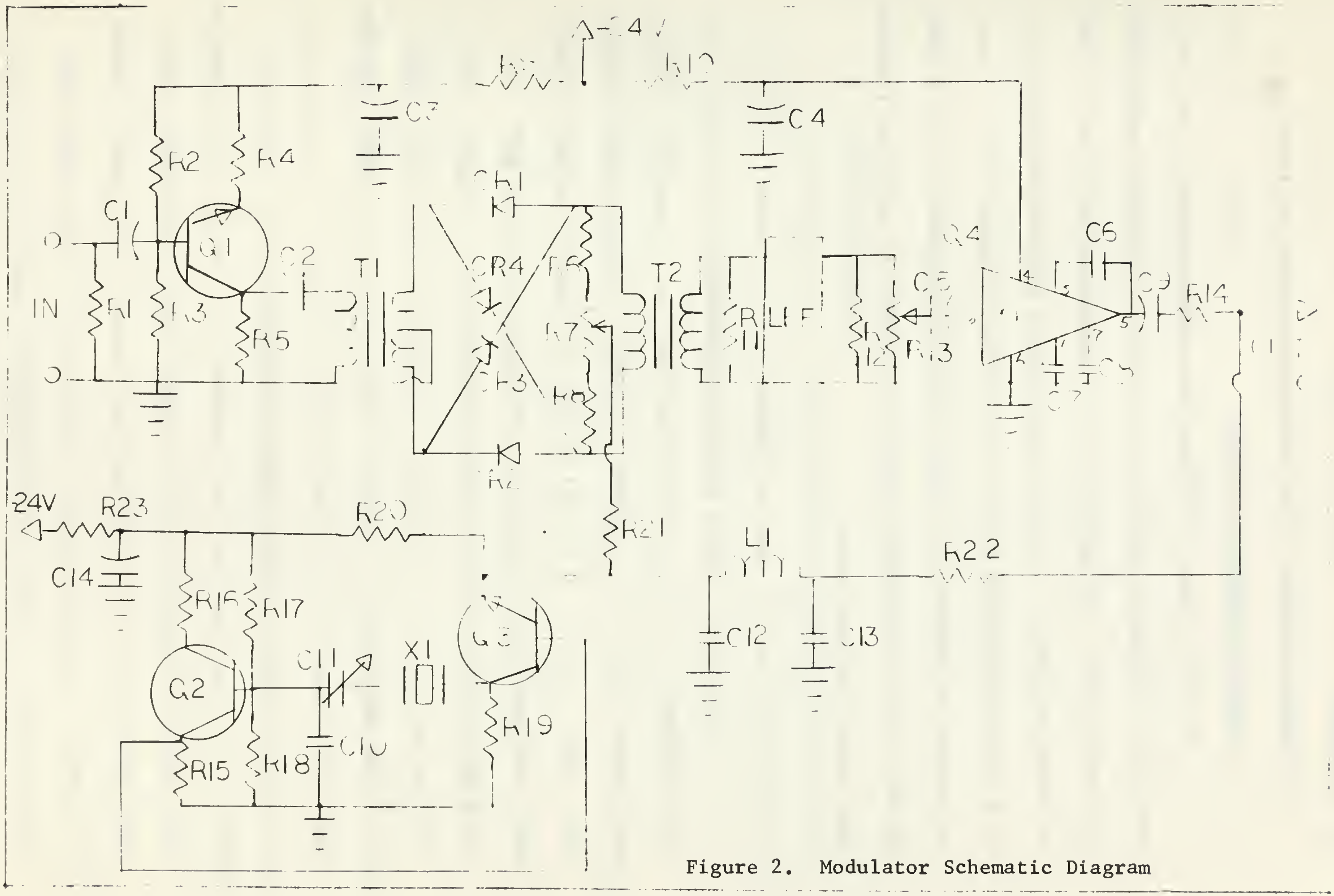


Figure 2. Modulator Schematic Diagram

B. MODULATOR

A schematic diagram of the modulator is shown in Figure 2.

1. Impedance-Matching Amplifier

The input stage of the modulator, consisting of Q1 and its associated components, provides the desired terminating impedance for the input signal. The amplifier has a voltage gain of about 2.5 and is biased at 4 ma. It is very linear due to emitter feedback. R1 is a 75-ohm terminating resistor for a 75-ohm coaxial line. It can be changed to terminate any other input impedance less than 600 ohms. R9 and C3 form a decoupling network to isolate the amplifier from the power supply and vice versa.

TABLE I

PARTS LIST FOR MODULATOR

R1	75 ohm	R9	100 ohm	R17	8.2K
R2	4.7K	R10	100 ohm	R18	100K
R3	50K	R11	1K	R19	1.8K
R4	240 ohm	R12	820 ohm	R20	1K
R5	3K	R13	2.5 K pot.	R21	1K
R6	1K	R14	75 ohm	R22	10K
R7	100 ohm pot.	R15	18K	R23	20 ohm
R8	1K	R16	1K		

All resistor 5%, ½watt

C1	1	C6	75 pf	C11	1-25 pf
C2	10	C7	10	C12	.006
C3	50	C8	1	C13	.0047
C4	50	C9	10	C14	50
C5	1	C10	200 pf		

TABLE I Continued

All capacitors microfarads, 25V, except as noted

Q1-Q3	2N3704
Q4	uA716 integrated circuit
CR1-CR4	1N34A
L1	2 millihenries
T1,T2	Triade SP 67, 1:1
X1	52 kHz quartz crystal
LPF	LASELCO 1352401 (47.7 kHz)

2. Balanced Modulator

The modulator is a ring modulator consisting of T1-2, CR1-4, and R6-8. Theoretically the output contains the sum and difference of the carrier and modulating frequencies and only odd harmonics of the carrier frequency. The sum frequency is the upper sideband and the difference frequency is the lower sideband. The carrier and modulating frequencies are suppressed. [4]

The carrier frequency is fed through R7 and acts as a switch which alternately switches the direction of the modulating current through the primary of T2 at the carrier rate. A positive carrier voltage turns on CR1 and CR2 and a negative voltage turns on the other two diodes. The incoming signal flows through the diodes that are conducting. The carrier is much larger than the signal and therefore controls the diodes. If the diodes are identical and if the secondary of T1 is perfectly center-tapped, R7 can be adjusted such that the carrier voltage drop across the primary of T2 is zero therefore no carrier will appear in the output. In a practical case, however, some carrier leak is always present and R7 is adjusted to minimize it.

Figure 3. Low-Pass Filter Frequency Response

Figure 4. Mixer Impedance Characteristics

3. Low-Pass Filter

The low-pass filter is a nine-pole elliptical filter with less than 0.5 db of ripple in the pass band of 0-47.7 kHz. The attenuation of frequencies above the pass band is greater than 55 db.. The frequency response of the filter is shown in figure 3. The input and output impedance of the filter is 600 ohms and it must be properly matched to obtain minimum ripple. The parallel combination of R12 and R13 provides a 620-ohm load for the output of the LPF. The input was matched by first determining the open-circuit output impedance of T2 as a function of R5 and then choosing R5 and R11 to provide a 600-ohm load. Figure 4 shows a plot of the output impedance of T2 as a function of R5. The output impedance of the mixer is 1500 ohms when R5 is 3000 ohms, and R11 was chosen as 1 K to provide a parallel combination of 600 ohms.

4. Output Amplifier

The output amplifier raises the signal level of the selected sideband to the required level of -15 dbm. The amplifier is a Fairchild uA716 integrated circuit designed especially for telephone use. The characteristics of the uA716 are shown in Appendix A. The uA716 can be configured to have a fixed voltage gain of 10, 20, 100 or 200. The output amplifier of the demodulator is configured to have a fixed gain of 100 which is sufficient to provide an overall gain in the modulator of 31.5 db. The output level can be adjusted by R13. The output impedance of the uA716 is approximately 1 ohm. A 75-ohm resistor is placed in series with the output of the uA716 to provide the proper source impedance to match a 75-ohm line. The series resistor is also required to develop the injected oscillator signal.

C6 is a frequency-compensation capacitor which prevents the amplifier from oscillating. C7 and C8 decouple pins 1 and 7 respectively which decreases the internal feedback to provide a fixed gain of 100. R10 and C4 form a decoupling network which isolates the amplifier from the power supply. The uA716 draws about 20 ma of current at 24 volts.

5. Oscillator

The oscillator consisting of Q2, Q3, and associated components generates the carrier frequency. The oscillator is essentially a crystal-controlled multivibrator and the output is almost a square wave. X1 is a 52-kHz quartz crystal which is the frequency-controlling device. The actual natural resonant frequency of the crystal is 51,997 Hertz. The series capacitor, C11, is adjusted to pull the frequency to 52 kHz. The purpose of C10 is to provide a low-impedance path for high frequencies and prevent spurious oscillations. R23 and C14 decouple the oscillator from the power supply. R21 is selected to lower the carrier level to about 1 volt rms for injection into the modulator.

6. Carrier Insertion Network

R22 and R14 form a voltage-divider network which lowers the level of the carrier to a value 10 db below test tone for insertion onto the output line. If R14 is changed to match a different line impedance, then R22 must also be changed to keep the insertion level constant. The bandpass filter, consisting of L1, C12, and C13, filters out the higher-order harmonics of the square-wave output of the oscillator and provides a clean sinewave for insertion onto the output line.

The inserted carrier serves as a pilot signal which is used to synchornize the demodulator.

C. DEMODULATOR

The demodulator translates the received signals back to their original position in the frequency spectrum. A schematic diagram of the demodulator is shown in figure 5.

1. Impedance-Matching Amplifier

The input stage consisting of Q1 and associated components provides the proper terminating impedance for the input line and also provides a 600-ohm source impedance to match the input to the low-pass filter. R1 is selected to provide the required terminating impedance and R5 is a 620-ohm resistor which provides the proper match for the low-pass filter. The gain of the input stage is about unity for lower-sideband frequencies. The reason for this is to keep the signals small to prevent distortion in the mixer. The output stage has sufficient gain to make up any losses. The input stage is very linear due to emitter feedback.

2. Low-Pass Filter

The low-pass filter, which is identical to the one used in the modulator, rejects the upper sideband and allows the lower sideband to pass into the mixer. The input to the low-pass filter is matched by R5. The output is matched by choosing the proper values for R6, R10, and R11 with the aid of the curve in Figure 4. In this case the series combination of R10 and R11 provides the load for the mixer.

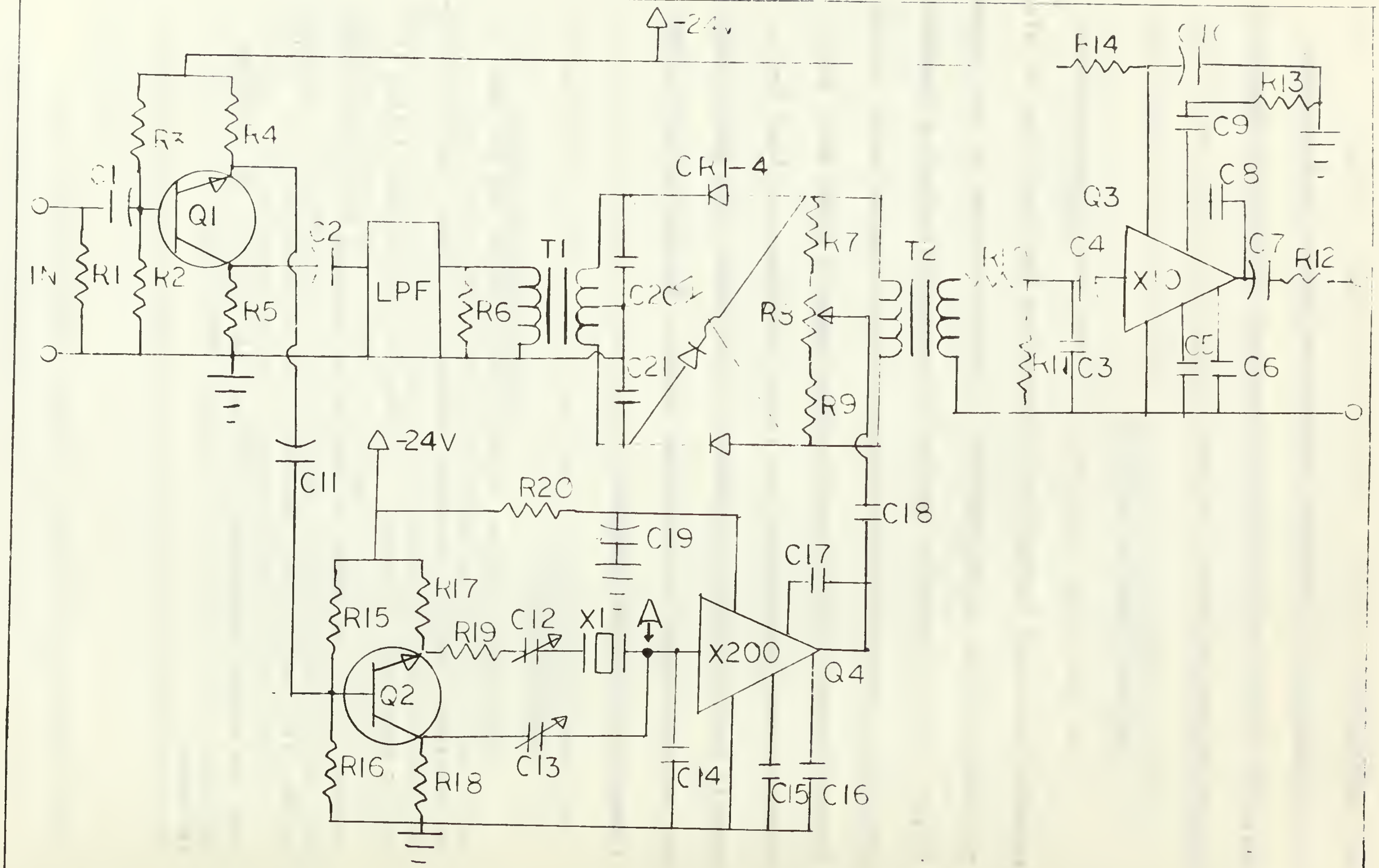


Figure 5. Demodulator Schematic Diagram

TABLE II

PARTS LIST FOR DEMODULATOR

R1	75 ohm	R8	100 ohm pot.	R15	8.2K
R2	24K	R9	1K	R16	27K
R3	5K	R10	1K	R17	3.9K
R4	330 ohm	R11	2.2K	R18	3.9K
R5	620 ohm	R12	75 ohm	R19	3.9K
R6	2.2K	R13	75 ohm	R20	200 ohm
R7	1K	R14	20 ohm		

All resistors 5%, ½ watt

C1	1	C8	39pf	C15	1
C2	10	C9	75 pf	C16	1
C3	.0015	C10	50	C17	75 pf
C4	10	C11	10	C18	10
C5	10	C12	1-25 pf	C19	50
C6	10	C13	1-10 pf	C20	.0015
C7	10	C14	250 pf	C21	.0015

All capacitors microfarads except where noted

Q1, Q2	2N3704
Q3, Q4	uA716 integrated circuit
CR 1-4	1N34A
T1, T2	1:1
X1	52 kHz quartz crystal
LPF	LASELCO 1352401 (47.7kHz)

3. Balanced Demodulator

The demodulator is identical to the modulator with the exception of C20 and C21 which have been added to increase the carrier current through the diodes. For proper mixing action without distortion the carrier amplitude should be at least ten times larger than that of the incoming signal. [5] The carrier level out of Q4 is 3 volts peak-to-peak and the nominal value of the input signal is 0.137 volts peak-to-peak, providing a 22:1 mixing ratio. But if all five incoming signals happen to be in phase at a particular instant the mixing ratio would be only about 5:1. Fortunately, as more channels are added the signal begins to approach the characteristics of white noise and the probability of all signals being in phase at once is small (for example, the average power contained in 12 voice channels is only 3.3 db greater than that in one channel). [6]

The capacitors, C20 and C21, are in parallel to the carrier but they are in series to the incoming signal; therefore, they offer a higher impedance to the incoming signal and do not greatly attenuate it.

4. Output Amplifier

The output amplifier is a uA716, operating with a voltage gain of 10, which is more than adequate to make up the losses in the demodulator and provide an overall gain of unity. The resistors R10 and R11 provides the load on T2 which is reflected back to the low-pass filter. C3 is used to suppress the upper sidebands. Actually, there should be another low-pass filter between T2 and Q3, but only two filters were available at the time the design was made. The combination of C8, C9, and R13 forms a frequency-compensation network recommended by the manufacturer. R14 and C10 decouple the amplifier from the power supply. R12 is chosen to provide a source impedance to match the output line. The uA716 is well adapted for this application. It will produce output voltage swings up to 12 volts p-p without distortion, and can drive almost any output impedance since the output impedance of the device is almost zero (see Appendix A).

5. <u>Carrier Pick-off Filter/Amplifier</u>

The incoming signal is taken from the emitter of Q1 and applied to a phase splitter consisting of Q2 and R15-19. [7] The resistors R17 and R18 are identical; therefore, the outputs at the emitter and collector of Q2 are equal in amplitude and opposite in phase. These two outputs are applied through two separate paths to point A where they are combined and cancel each other at frequencies other than the resonant frequency of the crystal. At the carrier frequency the crystal offers a low impedance and the two signals do not cancel, leaving an input to the amplifier at the carrier frequency. The crystal filter is tuned to the incoming carrier frequency by the series capacitor C12. To understand the cancelling action, assume an incoming signal different from the carrier frequency. The impedance looking from point A toward the collector is just the capacitive reactance of C13 in series with R18. Looking back toward the emitter, the impedance is the capacitive reactance presented by the crystal and its holder in series with C12, and R19. If C13 is adjusted to exactly equal the series capacitance of C12 and X1, the two signals will be equal and opposite at point A and will, therefore, cancel each other.

Actually, the resistive component in the emitter branch is R19 in series with the parallel combination of R17 and the output impedance of the emitter, which is about 25 ohms. But the tolerance of the resistors is 5%, or 195 ohms for a 3.9K resistor; therefore, the difference in resistive components is insignificant compared to the possible variations due to the tolerance.

The carrier frequency is applied to a uA716 in the gain of 200 configuration and then applied directly to the mixer. The output of the amplifier is about 3 volts peak to peak.

The frequency response of the filter and amplifier is shown in figure 6. The attenuation at frequencies below the carrier frequency varies somewhat with frequency but is everywhere more than 25 db in the 28-48 KHz lower sideband range except for one point near 30 KHz where the attenuation sharply decreases to only 12 db. This phenomenon is only about 5 Hertz wide at the 3 db points and is probably due to a spurious resonant mode of the crystal [8]. At frequencies above the carrier frequency, the attenuation is almost a constant 24 db.

The gain of the carrier frequency through the filter and amplifier is 36 db. If the signal frequency present at the output is considered as noise, the maximum signal-to-noise ratio is 50 db, assuming the input signal to be -15 dbm and the carrier input to be -25 dbm.

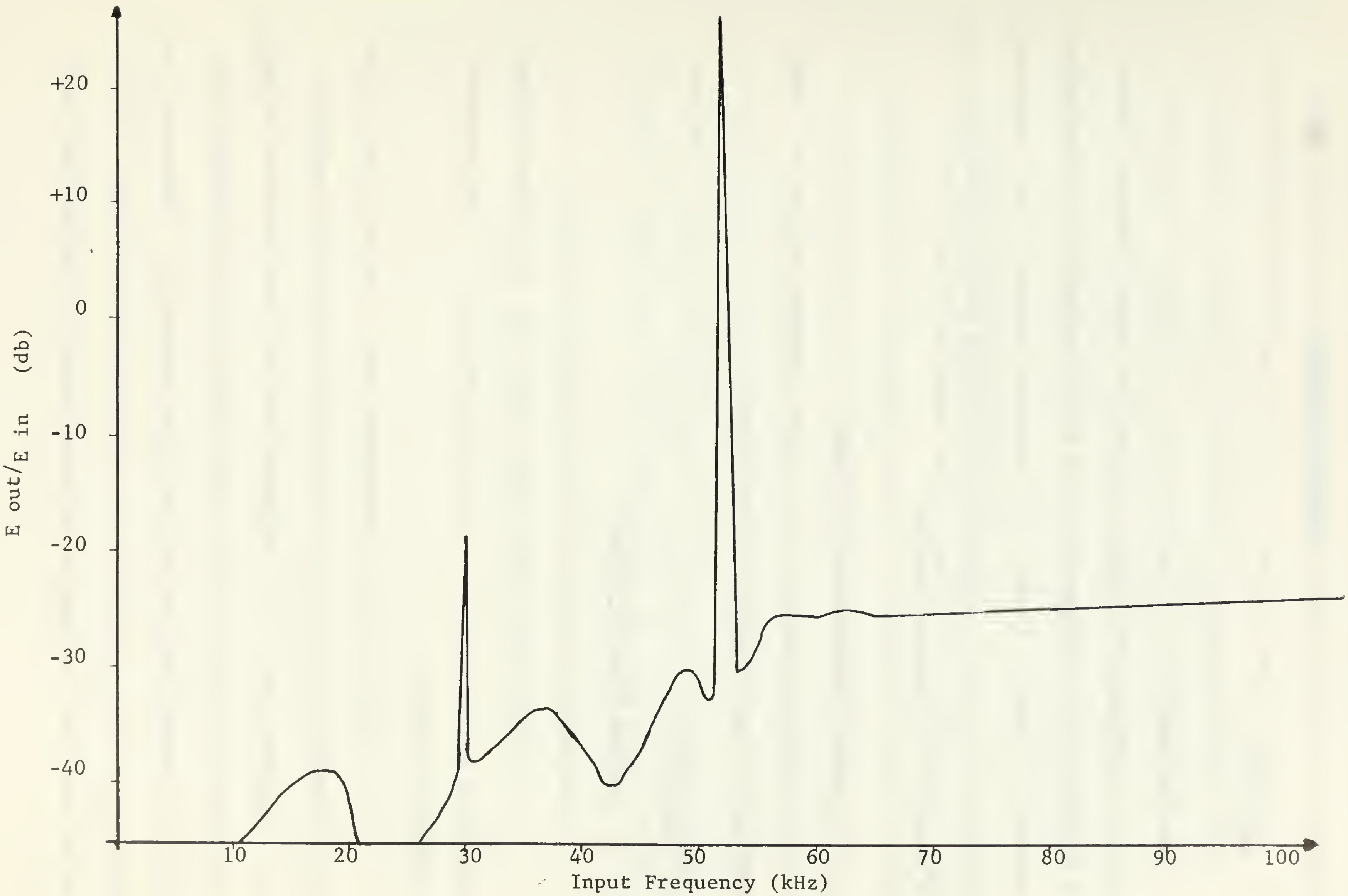


Figure 6. Carrier Pick-off Filter/Amplifier Frequency Response

V. TUNING PROCEDURE

A. MODULATOR

1. Oscillator

To properly tune the oscillator, the output frequency should be measured with a frequency counter and C11 adjusted until the output frequency is exactly 52 KHz. If a frequency counter is not immediately available, C11 can be adjusted to its midrange and the demodulator can be tuned to the inserted carrier.

2. Carrier Leakage from the Mixer

To tune the mixer, adjust R7 for a minimum reading on a VTVM connected across R11 with no signal input.

B. DEMODULATOR

1. Carrier Pick-off Filter

The carrier pick-off filter is tuned in two steps.

a. With normal signal input, adjust C12 for maximum carrier signal out of the amplifier. This tunes the filter to the inserted oscillator frequency.

b. With the input disconnected, apply a 66 Khz signal to the input terminals and adjust C13 for minimum output of the amplifier. It is important that C13 be adjusted when the input frequency is <u>above</u> the resonant frequency of the crystal because the attenuation of the filter is almost constant for the higher frequencies. This also gives the best over-all performance for frequencies below 52 KHz. If C13 is adjusted for a minimum in the above manner, it probably will not produce a minimum output at a randomly selected frequency below 52KHz, but if it is tuned to produce a minimum output at a frequency below 52KHz, the minimum at another frequency below 52KHz will probably be increased, and the minimum at frequencies above will amost surely be increased. The frequency of 66 Khz was selected because it is the center of the upper sideband; however, any frequency in the upper sideband range of 56-76 KHz would be satisfactory.

2. Carrier Leakage from the Mixer

The demodulator mixer is adjusted in the same manner as the modulator except that special care must be taken to ensure that there is no signal present at the input to the mixer while making the adjustment. This is because the input must be connected to the demodulator in order to have a carrier signal out of the amplifier. After the preceding precautions are taken, R8 is adjusted for minimum output across the output of T2. This measurement can be made with a VTVM.

VI. DESIGN ALTERNATIVES

A. MODULATOR

The design of the modulator is conventional with the exception of the carrier-insertion network. This network provides a high-impedance feedback path from the output back to the mixer which could possibly cause some distortion. This feedback path could be broken by placing an isolation transistor amplifier between R22 and the output (see figure 2). The added cost would have to be weighed against the improvement in performance.

B. DEMODULATOR

The carrier pick-off filter/amplifier is not conventional. The usual method for synchronizing such a system is to filter out the pilot carrier signal and use it to synchronize a separate oscillator. One of the advantages of that method is that voice communication is still possible in the unlikely event that the pilot signal alone is lost. In the normal case, however, if the pilot is lost, all signals are lost. Another advantage is that a separate oscillator would possibly introduce less noise into the mixer. The primary disadvantage of the separate oscillator is higher cost.

The 3-db bandwidth of the carrier pick-off filter is only 3 Hz. The small bandwidth is due to the extremely high Q of the quartz crystal. If the Q is defined as the center frequency divided by the bandwidth, then the Q of the filter is about 17,000. The narrow bandwidth could cause an alignment problem if either the oscillator or the filter were to drift due to temperature differences, aging, or other causes. The Q of the filter might be lowered, thereby increasing the bandwidth, by placing resistors in parallel with the output legs of the phase splitter.

The input signal level to the modulator is relatively large and there is the possibility that the mixer might become overloaded, especially if all five channels were operating at the same time or if there were several channels of digital information being sent. Therefore, an improvement in performance might be achieved by attenuating the input signal before applying it to the mixer and using more amplification in the output stage. The gain is available in the uA716 output amplifier at no extra cost.

VII. RESULTS AND CONCLUSIONS

A. FREQUENCY RESPONSE

The overall frequency response of the translator is shown in Figure 7. The maximum ripply in the frequency range of interest (4-24 kHz) is less than 1 db. The gradual decrease in response at frequencies above 24 kHz is due to the capacitor (C3 in Figure 5) which was placed across the output of the demodulator to attenuate the upper sidebands. The minimum ripple obtainable is limited by the ripple in the response of the low-pass filters, which is less than 0.5 db. The reason the overall ripple is higher is because the filter in the demodulator was deliberately mismatched. It was found that the resistance required to match the filter loaded the mixer and caused an increase in harmonic distortion. It was decided that 1 db of ripple could be tolerated as a compromise between minimum ripple and minimum harmonic distortion.

B. DISTORTION

In general, any nonlinearity in an active device or in the external circuit elements connected to an active device, will lead to a distorted output waveform. If the input wave is a single sinusoid, the output wave will contain higher-order harmonic frequency components in addition to the input-frequency component. This is called harmonic distortion. If the input contains several sinusoidal components, the output, in general will contain not only higher-order harmonics of each input frequency but also frequencies equal to all sums and differences of the input frequencies and their integral multiples. This is called

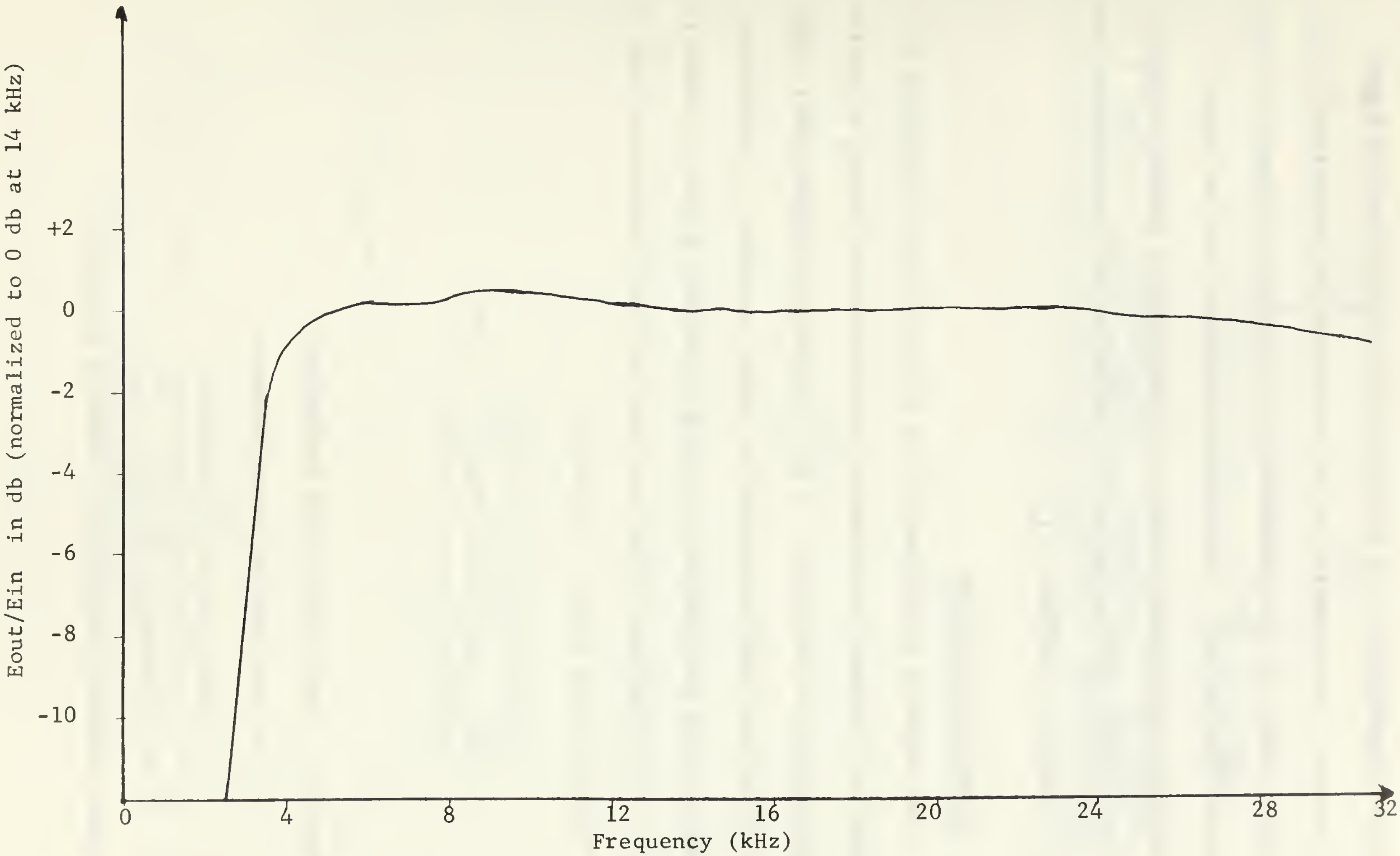


Figure 7. Overall Frequency Response of Translator

intermodulation distortion or crosstalk. In some applications the higher-order harmoics or the intermodulation components may be the desired output. For example, in the modulator and demodulator mixers the desired output is the intermodulation components between the input signals and the carrier. Intermodulation between any two input frequencies is undesirable, however. Distortion is present to some degree in all amplifiers, but may be controlled to an acceptable level by use of negative feedback. [4, 8]

1. Harmonic Distortion

The harmonic distortion of the translator was determined by measuring the frequency spectrum of the output when feeding a single-frequency signal to the input. [9] (A Hewlett-Packard model 302A wave analyzer was used to measure the output spectrum) A low-distortion input signal was obtained by using a low-pass filter as shown in Figure 8. (the low-pass filter was used only for frequencies below 10 kHz) The distortion factor is given by:

$$d = \frac{\sqrt{a_2^{\,2} + a_3^{\,2} + \ldots\ldots + a_n^{\,2}}}{a_1} \times 100$$

Where: a_1 = amplitude of the fundamental

a_2 = amplitude of second harmonic

a_n = amplitude of n^{th} harmonic

d = distortion (in %)

The overall harmonic distortion is shown in figure 8.

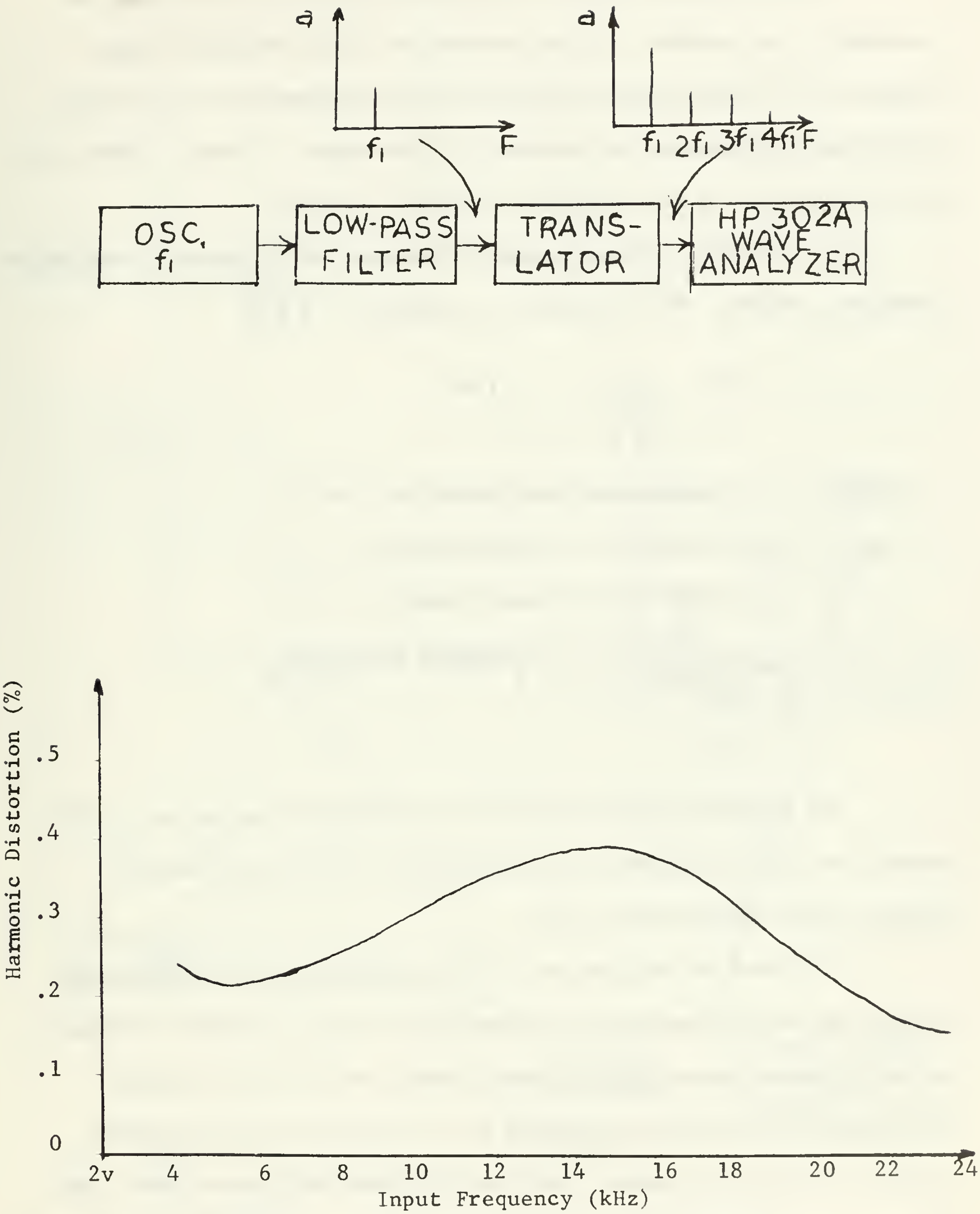


Figure 8. Harmonic Distortion Measurement

2. Intermodulation Distortion

The nonlinearity of the translator was also measured by the intermodulation method. In this method two equal-amplitude input signals are fed into the translator and the fundamentals and sum and difference frequencies are measured in the output. Figure 9 shows the setup for measuring the intermodulation distortion.

In the CCITT (International Telegraph and Telephone Consultative Committee) method, the distortion is given by: [9]

$$di = \frac{ad}{a_1+a_2} \times 100$$

Where: d_i = intermodulation distortion (in %)

a_1 = amplitude of input signal f_1

a_2 = amplitude of input signal f_2

a_d = amplitude of difference frequency signal $(f_2 - f_1)$

and $a_1 = a_2$

The intermodulation distortion was measured using the CCITT method, for many different combinations of f_1, f_2 and f_d and the maximum value observed was 0.14%.

The level of distortion that is tolerable in audio-frequency systems depends on the nature of the audio sounds involved, and upon a variety of psycho-acoustic factors. About one per cent harmonic distortion can usually be detected by a listening test but before harmonic distortion becomes intolerable it must be greater than about 10 per cent. Intermodulation as determined by the CCITT method can be detected at levels as low as a small fraction of 1 per cent in a

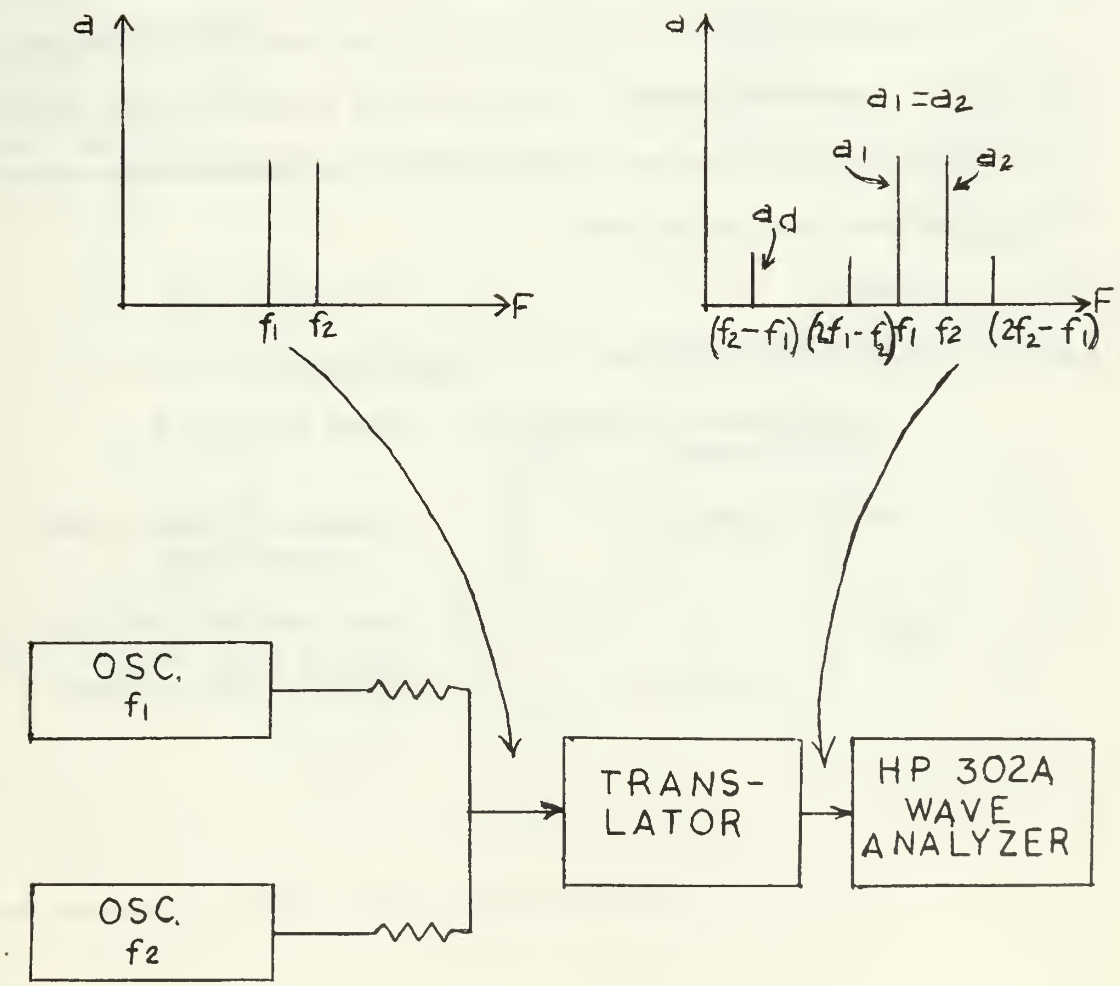


Figure 9. Intermodulation Distortion Measurement

listening test, and becomes objectionable at a value of 3 to 4 per cent when the difference frequency lies in the range 400-5000 Hertz, which is where the ear is most sensitive. [9] In general, the seriousness of audible intermodulation distortion in a telephone system depends on whether or not this crosstalk is intelligible. A lot of audible crosstalk can be tolerated if it is unintelligible, but if it is intelligible a small amount can be extremely annoying and distracting to the listener.

C. CONCLUSIONS

It is believed that the performance of the translator proposed in this thesis compares favorably with those now in general use, and that it is well suited for use on a small-capacity microwave link. The results achieved are as follows:

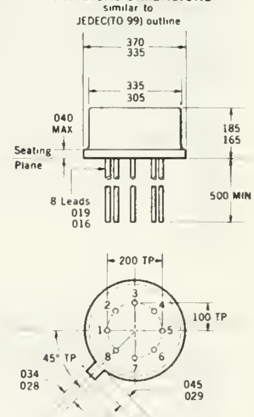
Ripple	Less than 1 db
Harmonic Distortion	Less than 0.4 %
Intermodulation Distortion (CCITT Method)	Less than 0.2 %
Power Consumption	Less than 2 watts for 5 channel group
Size	Each modulator or demodulator can be built on a 4" x 5" printed circuit board

GENERAL DESCRIPTION — The µA716 is a fixed-gain, medium power amplifier intended for use as a telephone system channel amplifier, headset amplifier or a general-purpose audio preamplifier. It provides medium output current capability, low distortion, excellent gain stability, and wide bandwidth. Fixed voltage gains of 10, 20, 100, and 200 are available by selecting external taps.

ABSOLUTE MAXIMUM RATINGS

Supply Voltage	27 V
Internal Power Dissipation (Note 1)	400 mW
Input Voltage	± 5 V
Peak Output Current ($T_A = 25°C$)	100 mA
Storage Temperature Range	− 65°C to + 150°C
Operating Temperature Range	− 55°C to + 125°C
Lead Temperature (soldering, 60 seconds)	300°C

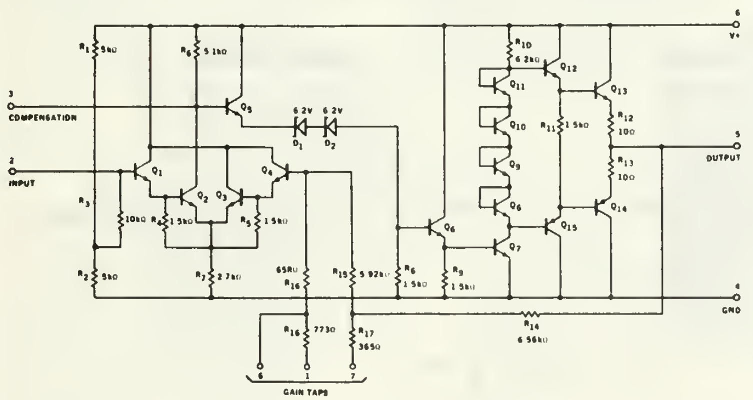
PHYSICAL DIMENSIONS
Similar to
JEDEC(TO 99) outline



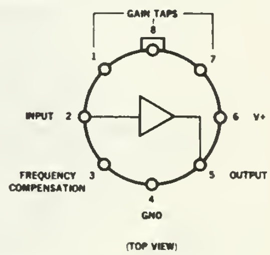
NOTES: Dimensions as per latest J 10 committee
All dimensions in inches
Leads are gold plated Kovar

ORDER PART NO. U5K7716312

SCHEMATIC DIAGRAM



CONNECTION DIAGRAM



NOTE: Pin 4 internally connected to case.

NOTE 1: Rating applies for case temperatures to +125°C; derate linearly at 8.4 mW/°C for ambient temperature above +110°C.

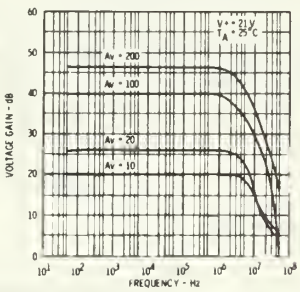
FAIRCHILD LINEAR INTEGRATED CIRCUITS μA716

ELECTRICAL CHARACTERISTICS ($-55°C \leq T_A \leq +125°C$, V+ = 21 V unless otherwise specified)

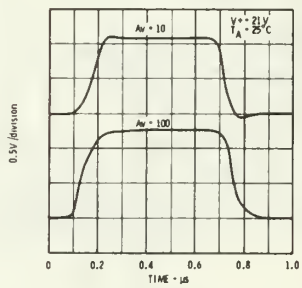
PARAMETER (see definitions)	CONDITIONS	MIN.	TYP.	MAX.	UNITS
Quiescent Power Consumption	$T_A = 25°C$		286	298	mW
	$T_A = 125°C$		244	256	mW
Total Harmonic Distortion	$f = 1$ kHz, $A_V = 10$, $P_O = 50$ mW, $R_L = 150\Omega$		0.01	0.05	%
	$f = 1$ kHz, $A_V = 100$, $P_O = 50$ mW, $R_L = 150\Omega$		0.10	0.50	%
Input Noise Voltage	$R_S = 600\Omega$, $T_A = 25°C$, $B_n = 16$ Hz to 150 kHz		8.0		μV_{rms}
Output Voltage Swing	$R_L = 150\Omega$	10	12		V p-p
	$R_L \geq 5\ k\Omega$	15	17		V p-p
Input Resistance		9.0	11		kΩ
Output Resistance			1.0		Ω
Voltage Gain					
10x	See Table 1	9.0	10	11	
20x	See Table 1	18	20	22	
100x	See Table 1	95	105	115	
200x	See Table 1	185	205	225	
Bandwidth	$T_A = 25°C$		2.0		MHz
Temperature Stability of Voltage Gain	$T_{ref} = 25°C$				
10x				±0.50	dB
20x				±0.50	dB
100x				±0.50	dB
200x				±0.65	dB

TYPICAL PERFORMANCE CURVES

VOLTAGE GAIN AS A FUNCTION OF FREQUENCY

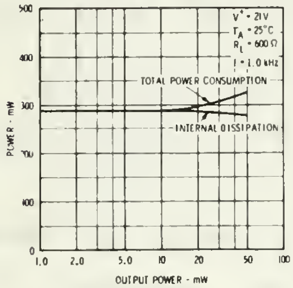


TRANSIENT RESPONSE

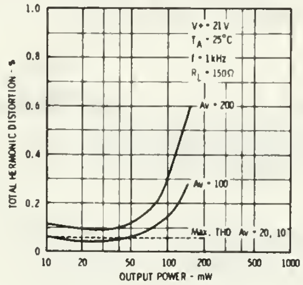


TYPICAL PERFORMANCE CURVES

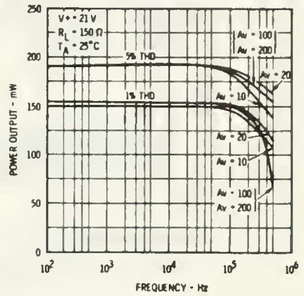
TOTAL POWER CONSUMPTION AND INTERNAL DEVICE DISSIPATION AS A FUNCTION OF OUTPUT POWER



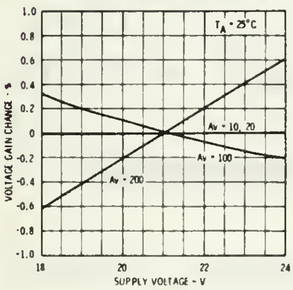
TOTAL HARMONIC DISTORTION AS A FUNCTION OF OUTPUT POWER



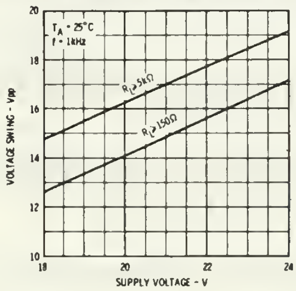
POWER OUTPUT AS A FUNCTION OF FREQUENCY 5% AND 1% TOTAL HARMONIC DISTORTION



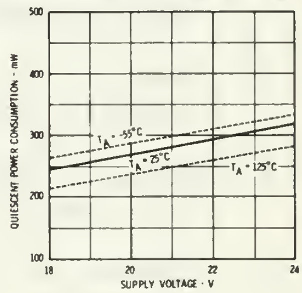
VOLTAGE GAIN AS A FUNCTION OF SUPPLY VOLTAGE



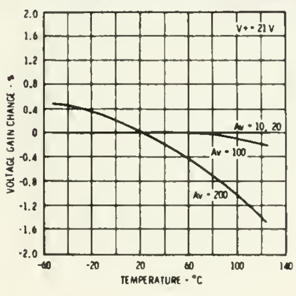
VOLTAGE SWING AS A FUNCTION OF SUPPLY VOLTAGE



QUIESCENT POWER CONSUMPTION AS A FUNCTION OF SUPPLY VOLTAGE



VOLTAGE GAIN AS A FUNCTION OF AMBIENT TEMPERATURE



CONNECTION DIAGRAM AND COMPONENT TABLE FOR AVAILABLE GAIN OPTIONS

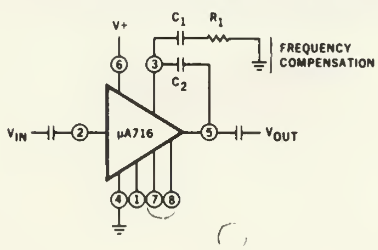


TABLE I

Voltage Gain	C_1	C_2	R_1	Decouple Pins:
10	68 pF	39 pF	75 Ω	1
20	50 pF	27 pF	75 Ω	8
100	None	3 pF	None	1, 7
200	None	3 pF	None	7, 8

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	ROLE	WT	ROLE	WT	ROLE	WT
Translator						
Multiplex						
Carrier-Telephone						